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COMPUTATION OF DELAY SPREAD USING 3D MEASUREMENTS

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Abstract – This paper proposes a new method for computing the delay spread (DS) of the mobile radio channel. The method utilizes a 3D measurement of the received signal at a given position and assumes knowledge of the receiver antenna radiation pattern. The main advantage of this method is that it allows testing of different antennas with only one propagation measurement. It also enables easy DS computation for different orientations of the receiver antenna. The paper describes the new method and investigates its performance by comparing the computed DS values with DS values obtained from measurements directly with the tested antennas. Three typical cellular handset antennas have been used. In addition the variation in DS is shown as function of azimuth rotation of the antenna.

I. INTRODUCTION

It is well known that the mobile radio channel is important in digital transmission systems because of the multipath propagation that takes place in the channel, generally making the channel both time-dispersive as well as time-varying [1]. The time-dispersion of the channel is an important issue in mobile radio system design. A system employing an equalizer can utilize the frequency diversity introduced by the channel, but only if the symbol duration of the system is small compared to the time-dispersion of the channel. On the other hand systems not using equalization of the channel can only operate satisfactorily if the time-dispersion is sufficiently small compared to the symbol duration.

The signal received by an antenna may be seen as a function of two processes. Firstly, the transmitted signal propagates through the mobile channel and is subject to scattering resulting in multiple rays arriving at the receiver location with different delays, amplitudes, phases, and directions. Secondly, the receiver antenna performs a weighted summation of the signals, with the weights

determined by the antenna pattern [2]. Thus, when the signal distribution at a given receiver location is known it is possible to compute the signal received by any antenna if the antenna pattern is known. This is a potentially powerful method for designing antennas, since propagation measurements only have to be carried out once and tests of antenna designs can be performed subsequently by simple calculations.

The DS of a channel is a well known measure of the time-dispersion of the received signals [1]. In this paper it is shown how the DS can be computed from knowledge of the 3D signal distribution and the receiver antenna pattern. The method is tested on measurements made in an urban micro cell propagation scenario where the 3D signal distribution is measured and the resulting DS is computed for three different antennas. The obtained results are compared to the DS computed from measurements directly using the three antennas.

II. MEASUREMENTS

The investigations were carried out using a dual channel correlation sounder for measuring the complex channel impulse responses (IRs) observed on the down-link between a base station and a mobile user in an urban environment. A center frequency of 1.89 GHz was used and the sounder had a bandwidth of approximately 20 MHz. The sounder has an instantaneous dynamic range of approximately 30 dB and an overall dynamic range of 80 dB.

The measurements were performed in an urban environment in the city of Aalborg with the transmitter on the roof of a twelve-story building which is higher than most of the surrounding buildings. The receiver was located approximately 700 m away inside a four-story office building and uses a dual polarized narrow-beam (about 30° beamwidth) horn antenna mounted on a po-

sitioning device controlled by step motors. The device, henceforth referred to as the *pedestal*, allows steering the horn antenna direction into a desired combination of azimuth and elevation angle. The direction of the horn antenna is controlled by an acquisition program collecting the measurements. For each position of the receiver a series of IR measurements was obtained with the antenna pointing in 72×15 directions, *i.e.*, all combinations of the azimuth angles $0^\circ, 5^\circ, 10^\circ, \dots, 355^\circ$ and the elevation angles $40^\circ, 45^\circ, 50^\circ, \dots, 110^\circ$, where the directions are given in usual spherical coordinates. IRs for both the θ - and the ϕ -polarizations were measured simultaneously. Each series of such measurements, subsequently referred to as a *record* of measurements, characterizes in 3D the actually received signals as seen by any antenna located at the same position. For these measurements it is assumed that no signals arrive from directions outside the measured set of directions and furthermore that the channel is essentially constant during the measurement duration of about 12 min. On each floor of the building three records of measurements were obtained inside the same room, where the measurement locations were separated by approximately 1.5 m.

For each location of the pedestal, a number of measurements were also made with a commercially available GSM-1800 handset. The handset has been modified to include a back mounted patch antenna in addition to the build-in helix/whip antenna. The patch antenna was designed for minimum radiation towards the user's head. The antennas are connected to the sounder via cables, and two antennas are measured at a time. The patch is used twice and denoted as either 'Patch(W)' or 'Patch(H)' depending on which antenna it is measured together with.

The rooms on the ground and the 2nd floor in which the measurements were made had windows towards the transmitter, and on the 1st floor the windows were in the opposite direction. The 3rd floor measurements were made in a hallway with no windows, but connected to rooms with windows. All the measurements were made in a non-line-of-sight situations.

III. DELAY SPREAD COMPUTATION

The received signal can be modeled as a linear convolution,

$$r(t) = \int h(\tau) s(t - \tau) d\tau$$

where the complex IR $h(\tau)$ is assumed to be time-invariant, $s(t)$ and $r(t)$ are the transmitted and received low-pass signals, respectively. Using that the signal re-

ceived by an antenna is a weighted integration of the signal distribution in space [2], it is straightforward to verify that the IR can be written as

$$h(\tau) = \oint E_\theta(\Omega) A_\theta(\Omega, \tau) + E_\phi(\Omega) A_\phi(\Omega, \tau) d\Omega \quad (1)$$

where E_θ and E_ϕ are the complex electric field components of the antenna, and A_θ and A_ϕ describe the signal distribution for the two polarizations as a function of the solid angle Ω . $A_\theta(\Omega, \tau)$ may be viewed as a directional IR for the θ -polarization, and similarly $A_\phi(\Omega, \tau)$ for the ϕ -polarization. The power delay profile (PDP) is defined as $P(\tau) = \mathcal{E}(|h(\tau)|^2)$ where $\mathcal{E}(\cdot)$ is the expectation operator. Using (1) the PDP is

$$P(\tau) = \oint |E_\theta(\Omega)|^2 R_\theta(\Omega, \tau) + |E_\phi(\Omega)|^2 R_\phi(\Omega, \tau) d\Omega \quad (2)$$

where it was used that

$$\begin{aligned} \mathcal{E}[A_\theta(\Omega, \tau) A_\theta^*(\Omega', \tau)] &= \mathcal{E}[|A_\theta(\Omega, \tau)|^2] \delta(\Omega - \Omega') \\ &= R_\theta(\Omega, \tau) \delta(\Omega - \Omega') \end{aligned}$$

and similarly for the ϕ -polarization. The cross-terms $\mathcal{E}[A_\theta(\Omega, \tau) A_\phi^*(\Omega', \tau)] = 0$. These assumptions can be verified given that the phases of $A_{\theta/\phi}$ are independent for different angles and furthermore independent of the amplitude.

The DS is computed from the PDP by the formula,

$$s = \left[\frac{1}{P_{\text{tot}}} \int (\tau - d)^2 P(\tau) d\tau \right]^{1/2} \quad (3)$$

where $P_{\text{tot}} = \int P(\tau) d\tau$ is the total received power and $d = \int \tau P(\tau) d\tau / P_{\text{tot}}$ is the expected delay.

IV. MEASUREMENT PROCESSING

For the handset measurements an instantaneous PDP is estimated from each IR of a measurement record, *i.e.*, $P_i(\tau) = |h_i(\tau)|^2$, where $h_i(\tau)$ is the i th IR of a record. In practice the measurements are noisy and this may have a significant influence on the estimated DS values. To reduce the influence of the noise a signal power 'window' is defined so that signals with a measured power outside the range $[P_{\text{max}} - P_{\text{dyn}}; P_{\text{max}}]$ is considered as noise and assigned a value of zero, where P_{max} is the maximum power of the measurement and P_{dyn} is the dynamic range, both in dB. For all the results given in this paper $P_{\text{dyn}} = 25$ dB, chosen after a visual inspection of the measured data.

Each of the PDPs is then used in (3), and thus 10^3 DS values are obtained from each of the handset measurement records. In the following the mean of these values is used as the DS for a given direction/antenna combination.

For the pedestal measurements the first step is to reduce the influence of the noise components in similar ways as described above. For a given antenna pattern the PDP is then computed using (2) from which the DS is obtained via (3). Using (2) the PDP depends on the orientation of the antenna in the environment, and thus also different DSs may be expected. Therefore, the DS is computed for the possible rotations (in azimuth) in 5° steps of the measured antenna pattern.

Since the DS computation from the pedestal measurements involves data from both polarizations, differences in branch powers, *i.e.*, imperfections in the measurement system, is corrected using back-to-back measurements before computing the PDP.

Although the horn antenna used for the pedestal measurements has a narrow beamwidth the measurements obtained with sampling intervals of only 5° in both azimuth and elevation angle cannot be considered independent. Because the beamwidth of the horn antenna is larger than 5° the received signal from one direction produces outputs at several directions. The horn antenna has two outputs, one for each polarization, with essentially no cross-coupling, and hence the output of one port can be written as [c.f. (1)]

$$v(\Lambda; \tau) = \oint E(\Lambda; \Omega) A(\Omega, \tau) d\Omega$$

where $v(\Lambda; \tau)$ is the IR measured in the direction $\Lambda = (\lambda, \mu)$, $E(\Lambda, \Omega)$ is the electric field pattern of the horn pointing in the Λ direction and A is the angular IR, both for either the θ - or the ϕ -polarization. Using the same assumptions as above the PDP is

$$P(\Lambda; \tau) = \mathcal{E}(|v(\Lambda; \tau)|^2) = \oint |E(\Lambda; \Omega)|^2 \mathcal{E}(|A(\Omega, \tau)|^2) d\Omega \quad (4)$$

Assuming $|A(\Omega, \tau)|^2$ is essentially constant within a local area, the expectation operator can be disregarded and only one pedestal measurement is necessary. Using this, Eq. 4 can be approximated as

$$P(\lambda, \mu; \tau) = \Delta_\theta \Delta_\phi \sum_\theta \sum_\phi |E(\lambda, \mu; \theta, \phi)|^2 |A(\theta, \phi, \tau)|^2 \sin(\theta)$$

where Δ_θ and Δ_ϕ are the angular sample intervals. Thus the measured IR in direction Λ may be seen as the output

of a 2D linear filter, where the filter IR is the determined by the horn antenna pattern. Since the channel is constant during a pedestal measurement A can be obtained from the horn antenna output by performing a deconvolution of the antenna pattern.

For a given value of τ the measurements $P(\lambda, \mu; \tau)$ for all combinations of values of λ and μ can be written as a matrix equation,

$$y = Bx + e \quad (5)$$

where y is a column vector of the $P(\lambda, \mu; \tau)$ values for the different measurement directions, B is a matrix containing the measured horn antenna pattern for all directions, and e is a noise vector. x is a column vector with the desired samples of A . From this model it is seen that A can be found by solving the matrix equation in (5) for x , and repeat this for each value of τ . In general the B matrix may be (nearly) singular and therefore Tikhonov regularization is used in this work [3]. In this method the solution \tilde{x} is found as

$$\tilde{x} = \sum_{i=0}^{N-1} \frac{\sigma_i}{\sigma_i + \alpha} \frac{u_i^T y}{\sigma_i} v_i \quad (6)$$

where σ_i 's are the N singular values of B , u_i and v_i the corresponding singular vectors, and $\alpha = \alpha_o \alpha_r$ is a regularization parameter. An optimum value of α_o is chosen according to the principle of generalized cross-validation [3]. $\alpha_r \neq 1$ is used to investigate the sensitivity to the α parameter. Due to noise, model inaccuracies *etc.*, the deconvolution may result in negative \tilde{x} elements. Since these elements represent power quantities this is non-physical. In this work the negative values are simply truncated to zero.

V. RESULTS

In Fig. 1–2 polar plots of the DS computed from pedestal measurements are shown for position no. 1 on each of the four floors. Each polar plot has curves for the four different antennas and the curves are shown as function of the azimuth angle. No deconvolution was applied before computation of these curves. In the same plots DS values obtained from handset measurements are shown with points. Comparing the two sets of DS values reveals a generally poor match between the individual DS values obtained with the handset and the corresponding values computed from pedestal measurements. However, there seems to be good match on a ‘coarse level’. From floor to floor the DS changes significantly for both sets of DS values.

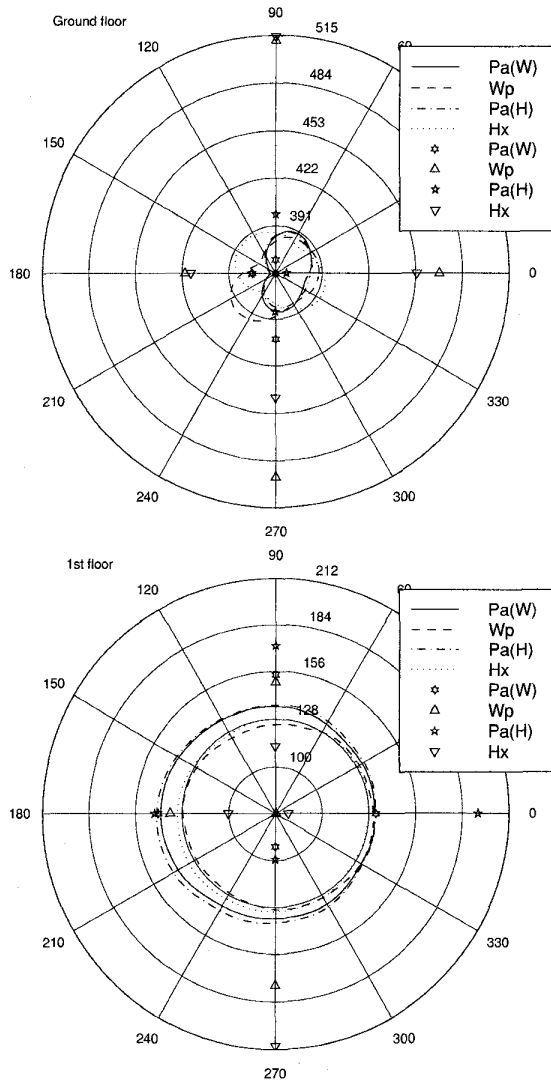


Figure 1: DS computed from pedestal measurements (curves) and DS computed from handset measurements (points). The values are in nano sec. Ground and 1st floor.

Floor	Patch(W)	Whip	Patch(H)	Helix
Gnd	2.7	-13.8	2.6	-15.1
1st	11.4	6.2	1.7	6.9
2nd	1.8	1.3	4.0	7.1
3rd	-23.3	21.7	-24.1	3.4

Table 1: Relative deviation in % of the handset DS from the pedestal DS. No deconvolution is used.

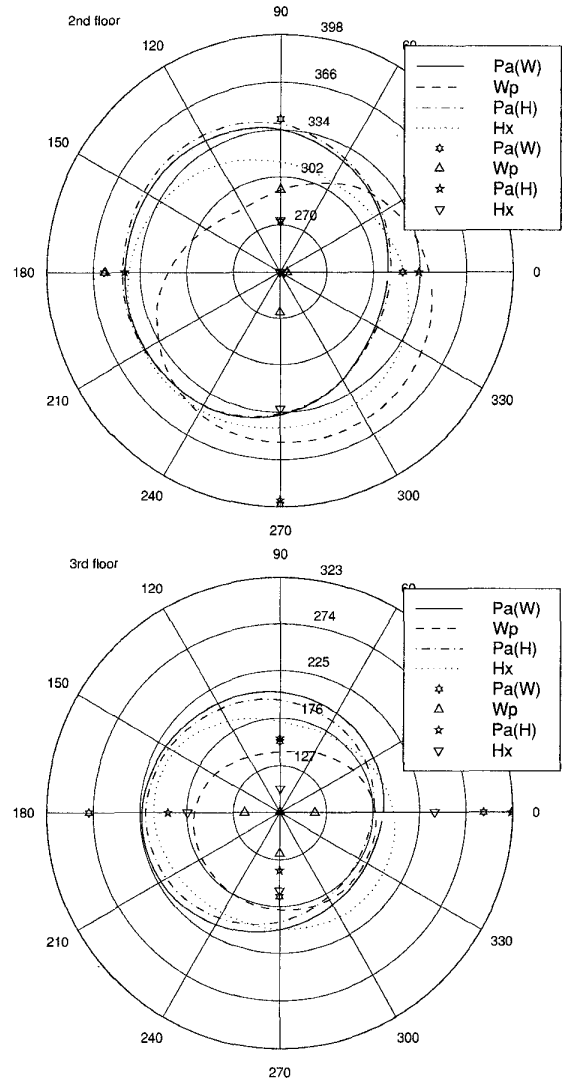


Figure 2: DS computed from pedestal measurements (curves) and DS computed from handset measurements (points). The values are in nano sec. 2nd and 3rd floor.

Floor	Patch(W)	Whip	Patch(H)	Helix
Gnd	-9.5	-35.2	-9.3	-32.9
1st	54.8	53.8	48.0	55.4
2nd	3.6	8.1	4.6	11.7
3rd	-37.4	16.8	-37.6	-7.9

Table 2: Relative deviation in % of the handset DS from the pedestal DS. Deconvolution is used with $\alpha_r = 1$.

Floor	Patch(W)	Whip	Patch(H)	Helix
Gnd	-17.3	-45.9	-16.5	-42.7
1st	24.4	22.3	15.4	24.5
2nd	-1.4	0.9	0.7	5.2
3rd	-44.7	9.6	-45.9	-14.6

Table 3: Relative deviation in % of the handset DS from the pedestal DS. Deconvolution is used with $\alpha_r = 5$.

The differences are investigated further in Tab. 1–3 where the mean of the quantity $100(s_p - s_h)/s_p$ is shown, where s_h is a DS obtained from a handset measurement, and s_p is the corresponding DS computed from a pedestal measurement. In other words the tables show the mean relative deviation, where the mean is over the different directions and positions. Tab. 1 shows that when no deconvolution is applied there is reasonably good match on the average with no apparent pattern. Applying deconvolution (Tab. 2–3) seems to make the predictions worse. The reason for this could be that the deconvolution as implemented is performed independently at the different delays and sometimes introduces non-existing signals to which the DS is sensitive.

The variation in the DS values as computed from the pedestal measurements is quantified in Tab. 4 where difference in minimum and maximum values are shown both in absolute numbers and relative to the minimum values. The min/max values are computed from each antenna/floor combination, i.e., for each curve in Fig. 1–2. From the table it is noticed that the variation in DS for rotation in azimuth is rather small and insignificant for systems such as GSM, where the symbol duration is $3.69 \mu\text{s}$. The small changes in DS is likely because of the types of antennas tested all of which have a fairly large beamwidth.

The relative deviations shown in Tab. 4 seem to depend on the floor, such that DSs on the ground and 1st floors show a smaller relative deviation than the DSs on the 2nd and 3rd floors, irrespective of the antenna type. A probable explanation for this is that on the topmost floors the power is more unevenly distributed than on the lower floors and therefore the antenna orientation is more important.

VI. CONCLUSION

In this paper a new method for computing the DS of a channel has been investigated. The proposed method uses a 3D measurement of the received signal and the radiation pattern of the receiver antenna, and has the advantage of allowing testing of different antenna designs with

Floor	Patch(W)		Whip		Patch(H)		Helix	
	Δ	R	Δ	R	Δ	R	Δ	R
Gnd	26	7	24	6	25	7	18	5
1st	9	7	8	6	12	9	5	4
2nd	33	11	67	23	33	10	39	13
3rd	40	22	42	30	46	26	40	24

Table 4: Difference $\Delta = s_{\max} - s_{\min}$ in maximum and minimum DS values and relative difference $R = \Delta/s_{\min}$ for pedestal position no. 1. The differences are shown in nano sec. and relative values are in %.

only one propagation measurement. With the purpose of verifying the method a number 3D measurements were made with a narrow beam horn antenna on a pedestal. Also measurements with a GSM-1800 handset equipped with three different antennas were made, and the radiation patterns of the antennas were measured. Comparing DSs computed from the two sets of measurements it was found that the two sets of DSs were in reasonable agreement on the average, as computed over different directions and positions. For the individual directions a precise match was not found which is believed to be because of inaccuracies in the measurements. Deconvolution of the antenna used for the 3D measurements was investigated and it was found not to be useful in this implementation, probably because of errors introduced by the deconvolution. However, the DS values agreed on the average without deconvolution being used. This is most likely because of the relatively large beamwidth of the antennas tested.

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